TITLE OF THE INVENTION

DIGITAL DATA PLAY BACK APPARATUS AND METHOD FOR PLAYING BACK DIGITAL DATA

CROSS-REFERENCE TO RELATED APPLICATIONS

5   This application is based upon and claims the benefit of priority from the prior Japanese Patent Application No. 2000-385342, filed December 19, 2000, the entire contents of which are incorporated herein by reference.

10   BACKGROUND OF THE INVENTION

The present invention relates to an improvement of a digital data play back apparatus and a method for playing back digital data which is optically, magnetically or optomagnetically recorded on a
15   recording medium such as, for example, a disc or the like.

As is already known, a play back signal processing system in a recording apparatus for optically, magnetically or optomagnetically recording data pattern
20   is generally constituted to obtain play back data by amplifying the play back signal read with a play back head from the recording medium and further equalizing a waveform of the play back signal with a waveform equalizer followed by inputting the play back signal to
25   a data detection device to identify data and decoding the signal to obtain reproduction data.

That is, a distortion generated as a result of the

passage of data through the recording channel is
corrected with the waveform equalizer and a signal
error detection ratio is suppressed within an allowance
scope with the result that a peak position of the

5   waveform is determined and a zero cross point is
determined through differentiation. In the PRML
(Partial Response Maximum Likelihood) signal processing
method in recent years, a method is used which detects
data as a sequence by making use of a correlation

10   before and after the signal.

Here, in order to detect data, an interval in
which data is recorded, namely, a channel clock
frequency, and phase information thereof are required.
This channel clock is recovered (restored) by using a

15   PLL (Phase Locked Loop) from the played back signal.

In the data recorded on the recording medium,
there is present at the initial portion of data a VFO
region in which a signal is recorded in a single data
repetition pattern repeated by integer times of the

20   clock for recording data for timing recovery. The peak
position of the VFO signal after the waveform
equalization, and the position at which the signal
crosses a specific threshold value are detected, and
the channel clock is recovered thereby enabling

25   synchronization to the frequency and phase of the
recorded data.

The following data is detected by using this

synchronized clock. However, because of the change in the rotational speed of discs or the like, the phase at the initial portion of data does not agree with the phase at the end portion of the data. Then, even during the play back of the data, the phase shift between the data detection timing and the play back clock by the PLL is feed back to the PLL thereby allowing the phase drift at the data portion to follow the clock.

Since the VFO region has a signal of a single frequency, a high quality phase signal can be obtained. However, since the data portion has a complicated waveform, the position at which the phase error can be detected is small, the quality thereof is lowered under the influence of the waveform interference. In the VFO region, the PLL is normally referred to as the "Acquisition mode" wherein the frequency is acquired at a high speed to follow the phase. However, in the data region, the PLL moves to the "trace mode" which is an operation in which the loop gain of the PLL is suppressed to a low level. In this trace mode, a mild frequency change like a rotational drifts can be followed, but a rapid change in bit unit by noises, waveform interference or the like cannot be followed.

In the PRML signal processing method, there are many cases in which the play back signal is quantized at a discrete time by using an A/D (Analogue/Digital)

converter prior to the inputting of the play back signal to the waveform equalizer. In the sampling clock given to the A/D converter, the channel clock is used which is recovered by using the PLL.

5   FIGS. 1A through 1H are views showing a signal of an optical disc apparatus in the case where the play back signal is not sampled in a discrete time as used many times in the PRML signal processing, but the threshold value is detected in a continuous time to

10   detect data.

The user data to be recorded is a data row which is a random combination of 0 and 1. In order to record the data on the recording medium, the number of 0's between 1 and 1 as shown in FIG. 1A is modulated

15   according to a run length limit (RLL) code. The data writing signal is converted into a signal (FIG. 1B) in which the ON and OFF of data (FIG. 1A) is toggled at the position of 1 in the case of mark edge recording. As a consequence, a mark (FIG. 1C) is recorded on the

20   recording medium. The play back signal (FIG. 1D) in the case where this mark (FIG. 1C) is read is allowed to pass through a differential circuit so that a differential waveform (FIG. 1E) can be obtained. The peak of the differential waveform (FIG. 1E) corresponds

25   to the position of 1 of the data which is modulated with the RLL code of the data (FIG. 1A).

The channel clock recovered in the PLL circuit by

using the signal of the VFO region may assume the state
shown in FIG. 1G. If so, this clock (FIG. 1G) can form
the data detection window (FIG. 1F). If the detection
window (FIG. 1F) has a differential-waveform peak of 1,
5 the detection data (FIG. 1H) can be acquired by
outputting "0." The detection data (FIG. 1H) is RLL
code data. Hence, the user data recorded can be
recovered by allowing the detection data (FIG. 1H) to
pass through the decoding circuit.

10 In the case where the phase of the play back
signal and the phase of the channel clock synchronize
each other, the peak of the differential signal
(FIG. 1E) is located in the center of the data
detection window (FIG. 1F). Consequently, although it
15 is easy to detect whether or not the peak is present in
the data detection window (FIG. 1F), the peak of the
differential signal (FIG. 1E) is shifted to the end of
the data detection window (FIG. 1F) along with the
shift of the phase, so that the peak ultimately enters
20 into the adjacent window thereby generating a detection
error.

Then, as shown in FIG. 2, the phase error between
the peak position of the differential waveform and the
channel clock is detected with the phase comparator.
25 And the error is converted into a voltage value with a
charge pump to control the voltage control oscillator
(VCO) as a result of the conversion. As a consequence,

the clock phases which are recovered with the play back signal and the PLL can be allowed to synchronize each other.

In such method, a detection error is likely to be generated when the mark edge is not raised with the time in the data detection window. When the recording density is heightened, the rise of the edge becomes mild under the influence of the frequency response characteristics of the recording medium. In order to correct the rise to more abrupt rise, it is required to amplify the high frequency component characteristic with a waveform equalizer. However, since the noise component in a high frequency area is also raised in this correction, the S/N is deteriorated. Thus, in this signal processing method, there is a limit in the heightening of the density.

One method of overcoming this problem is a PRML signal processing method. In the PRML, since the influence of the response waveform to a certain bit is allowed to be exerted upon the adjacent bit or a plurality of bits, data can be recorded in a high density without emphasizing the high frequency component of the response of signal processing channel.

Instead of this, the degree of influence upon a plurality of bits is required to be controlled to a value determined with the PR class. Since the degree of the mutual interference between bits is controlled,

the waveform having a mild change which cannot be detected in the threshold value is optimally detected (an ML detection) in a sequence of waveforms so that data can be detected without errors.

5   FIGS. 3A through 3F are views showing how the data row (FIG. 3A) which is the same as FIG. 1A will be represented with in the PRML method. FIG. 3C is a view showing a bit row wherein the case in which the writing signal (FIG. 3B) for each channel clock is turned on is 10 set to 1 while the case in which the signal is turned off is set to 0. Now, suppose that the PR class is (1, 2, 2, 1) as shown in FIG. 3D.

In this PR equalization, it is intended to show that a response waveform played back from the channel 15 in the case where the writing bit is 1 will be as shown in FIG. 3D. As shown in FIG. 3E, the play back waveform corresponding to the writing bit row (FIG. 3C) is represented as a superposition of the response waveform of each bit, so that the waveform shown in 20 FIG. 3F can be obtained.

Thus, in the case of the PR equalization, what is controlled with the PR equalizer and is used for the input of the ML detector is the amplitude value at the time of sampling. Consequently, in the PRML signal 25 processing, the play-back signal is converted into an amplitude value system quantified at the discrete time for each channel clock with the A/D converter before

the play back signal is input to the equalizer. After
that the play back signal is processed with the digital
circuit in almost all the cases.

In the PLL phase error detector used in the PRML
signal processing system optical disc apparatus, an
output signal of the equalizer is made discrete in the
direction of time. Unlike the method explained by
using FIGS. 1A through 1H, the data phase and the phase
of the channel clock to be recovered with the PLL
cannot be directly compared with each other. It is
necessary to convert a shift in the sampled amplitude
value into a phase error.

For example, USP No. 4,890,299 discloses a method
for converting a shift in the amplitude value into a
phase error and a structure thereof. A calculation is
made as to what degree of timing phase gradient has
either in the plus or minus direction from the current
amplitude value, amplitude values before one or two
samples, and an ideal equalization amplitude
corresponding to the values.

However, what is disclosed here is a case in which
an ideal value of the play back waveform after the PR
equalization becomes three levels. When the recording
density becomes high, the frequency response of the
medium becomes relatively insufficient so that the PR
class needs to be raised. However, since the points of
samples in which waveforms mutually interferes with

each other increases in these classes, the amplitude
level increase after the equalization of the waveform
with the result that the phase error cannot be
detected in the method in which the three levels are

5   assumed.

How many levels the waveform value is divided into
after the PR equalization depends upon the PR class and
the modulation method used. As in the examples of
FIGS. 3A to 3F, the waveform value is divided into five

10   level (0, 1, 3, 5 and 6) if the levels (1, 2, 2, 1) are
used as the PR class and if PLL (2, 7), wherein the
minimum number (d constraint) of 0s existing between
"1" and "1" is limited to 2, is used as modulation
method.

15   Even if the same PR class, i.e., the levels (1, 2,
2, 1) is used, the waveform value will be divided into
seven levels (0, 1, 2, 3, 4, 5 and 6) in the case where
the d constraint is 0 or the PLL (1, 7), wherein the d
restriction is limited to 1, is used as modulation

20   method. This holds true when the maximum number (k
constraint) of 0s existing between "1" and "1" is 4 or
more.

In this manner, in the case of the class in which
the level of the waveform value after the PR

25   equalization is 5 through 7, it is impossible to
calculate the timing phase gradient from the amplitude
value of the waveform in the method disclosed in USP

No. 4, 890,299.

Incidentally, with respect to this problem, as shown in Jpn. Pat. Appln. KOKAI Publication No. 2000-195191 filed by the same applicant of the

5 present invention, a countermeasures technique is already filed which facilitates a phase synchronization between data and a block by detecting a timing phase gradient from the amplitude value sampled at a discrete time also in the PR class in which the amplitude value

10 after equalization becomes 5 through 7 levels. However, in the current situation in which technical development is briskly made, there can be seen a tendency that a countermeasures having a different structure is also desired.

15 BRIEF SUMMARY OF THE INVENTION

The present invention has been made in view of the above circumstances, and an object of the invention is to provide a digital data play back apparatus and a method for playing back digital data which enables a

20 synchronous control of the recovered channel clock with the data played back in the user data region in the case of the PR class in which the level of the waveform value after the PR equalization becomes 5 through 7.

The digital data play back apparatus according to

25 the present invention comprises:

a conversion portion for converting a signal read from a recording medium into digital data by sampling

the signal in a clock frequency output from a variable
frequency oscillation portion;

an equalizing portion for equalizing digital data
output from this conversion portion into a waveform
which agrees with a partial response class on the basis
of a clock output from a variable frequency oscillation
portion;

a data detection portion for detecting data on the
basis of a clock output from the variable frequency
oscillation portion with respect to the signal waveform
output from this equalizing portion;

a timing phase gradient detection portion for
detecting a gradient of a phase error between a signal
waveform output from the equalization portion and a
clock output from the variable frequency oscillation
portion; and

a control portion for controlling the oscillation
frequency of the variable frequency oscillation portion
on the basis of the timing phase gradient detected with
the timing phase gradient detection portion.

Furthermore, the method for playing back digital
data according to the present invention, comprising:

a first step of converting a signal read from the
recording medium into digital data by sampling the
signal in a clock frequency output from a variable
frequency oscillation portion;

a second step of equalizing the digital data

output at the first step into a waveform which agrees
with a partial response class on the basis of a clock
output from the variable frequency oscillation portion;

a third step of detecting data on the basis of the

5   clock output from the variable frequency oscillation
portion with respect to the signal waveform output at
the second step

a fourth step of detecting a gradient of a phase
error between a signal waveform output at the second

10 step and a clock output from the variable frequency
oscillation portion; and

a fifth step of controlling the oscillation
frequency of the variable frequency oscillation portion
on the basis of the timing phase gradient detected at

15 the fourth step.

According to the above structure and the above
method, since the timing phase gradient can be detected
from the amplitude value sampled at the discrete time
also in the PR class in which the amplitude level

20 after equalization becomes a value of 5 through 7, a
synchronous control of the data played back in the user
data region and the recovered channel is enabled with
high reliability also in the case of the PR class in
which the amplitude level after equalization becomes 5

25 through 7.

Additional objects and advantages of the invention
will be set forth in the description which follows, and

in part will be obvious from the description, or may be

learned by practice of the invention. The objects and

advantages of the invention may be realized and

obtained by means of the instrumentalities and

5   combinations particularly pointed out hereinafter.

BRIEF DESCRIPTION OF THE SEVERAL VIEWS OF THE DRAWING

The accompanying drawings, which are incorporated

in and constitute a part of the specification,

illustrate presently preferred embodiments of the

10   invention, and together with the general description

given above and the detailed description of the

preferred embodiments given below, serve to explain the

principles of the invention.

FIGS. 1A through 1H are views for explaining a

15   timing relation between a writing signal and a read-out

signal with respect to an optical disc apparatus

respectively;

FIG. 2 is a block structure diagram shown for

explaining a PLL for allowing a phase of a play back

20   signal and a phase of a channel clock in the optical

disc apparatus to agree with each other;

FIGS. 3A through 3F are views for explaining a

timing relation between a writing signal and a read-out

signal on the basis of the PRML method in the optical

25   disc apparatus respectively;

FIG. 4 is a block structure diagram for explaining

an outline of the optical disc apparatus, the view

showing a first embodiment of the present invention;

FIG. 5 is a block structure diagram shown for explaining another example in which the read channel clock and the play back signal are allowed to synchronize with each other in the first embodiment;

FIG. 6 is a block structure diagram for explaining the details of a timing phase gradient detection circuit in the first embodiment;

FIGS. 7A through 7F are views for explaining a VFO pattern for detecting a timing phase gradient in the first embodiment respectively;

FIGS. 8A and 8B are views for specifically explaining an operation of detecting the timing phase gradient in the first embodiment respectively;

FIGS. 9A and 9B are views for specifically explaining an operation of detecting the timing phase gradient in the first embodiment respectively;

FIG. 10 is a flowchart shown for explaining in summary an operation for detecting the timing phase gradient in the first embodiment;

FIG. 11 is a block structure diagram for explaining the details of the timing phase gradient detection circuit, the view showing a second embodiment of the present invention;

FIG. 12A and FIG. 12B are views for specifically explaining an operation of detecting a timing phase gradient in the second embodiment, respectively;

FIG. 13 is a view shown for specifically explaining an operation of detecting the timing phase gradient in the second embodiment.

FIG. 14 is a flowchart shown for explaining in summary an operation of detecting the timing phase gradient in the second embodiment;

FIGS. 15A through 15C are state transition diagrams for explaining how the amplitude value of the PR equalization signal changes with the data row in the second embodiment respectively;

FIGS. 16A and 16B are views shown for explaining an amplitude pattern which is obtained at the rise of the waveform in the case of d = 0, 1, 2 in the second embodiment respectively;

FIG. 17 is a view for explaining an amplitude pattern which can be taken at the rise of the waveform in the case of d = 0, 1, 2 in the second embodiment;

FIG. 18 is a block structure diagram shown for explaining the details of the timing phase gradient detection circuit, the view showing a third embodiment of the present invention;

FIGS. 19A through 19F are views shown for explaining a state of transition of the waveform in the case in which g(n-1) is other than 0 in the third embodiment respectively;

FIGS. 20A through 20F are views for explaining the state of transition of the waveform in the case where

g(n-1) is 0 in the third embodiment respectively;

FIGS. 21A through 21F are views for specifically explaining an operation of detecting a timing phase gradient in the third embodiment respectively;

5   FIG. 22 is a flowchart for explaining in summary the operation of detecting the timing phase gradient in the third embodiment;

FIG. 23 is a block structure diagram shown for explaining in detail the timing phase gradient

10   detection circuit, the view showing a fourth embodiment of the present invention;

FIGS. 24A through 24F are views for specifically explaining the operation of detecting the timing phase gradient in the fourth embodiment respectively;

15   FIG. 25 is a flowchart for explaining in summary the operation of detecting the timing phase gradient in the fourth embodiment;

FIG. 26 is a block structure diagram shown for explaining the details of the phase error detection

20   circuit, the view showing a fifth embodiment of the present invention; and

FIG. 27 is a flowchart for explaining in summary the operation of detecting the timing phase gradient in the fifth embodiment.

25   DETAILED DESCRIPTION OF THE INVENTION

Hereinafter, a first embodiment of the present invention will be explained in detail by referring to

the drawings. FIG. 4 is a view showing an essential portion of an optical disc apparatus which will be explained in the first embodiment. That is, the optical disc apparatus primarily comprises a rewritable

5   disc 1 which is a recording medium, an optical head 3 for writing and reading data to the optical disc 1, a data play-back system, a data recording system, a drive controller 12 and an interface 13.

The optical disc 1 is rotated and driven with a

10   spindle motor 2. The optical head 3 is moved and adjusted with a servomotor 4, so that laser light is applied to the optical disc 1 with the drive by the laser driver 15 thereby optically recording and playing back information. The spindle motor 2 and the

15   servomotor 4 are driven and controlled with a drive controller 12 via a drive control circuit 14.

The data recording system has a laser driver 15 and a modulation circuit 16. The modulation circuit 16 conducts coding processing for converting record data

20   sent from the drive controller 12 into a predetermined code bit row. The laser driver 15 drives the optical head 3 so as to record on the optical disc 1 a mark which follows a code bit row output from the modulation circuit 16.

25   The data play back system has a preamplifier 5, a VGA (variable gain amplifier) 6, an A/D conversion circuit 7, an equalizer 8, a data detection circuit 10

and a decoder 11. The preamplifier 5 and the VGA 6 amplify the playback signal read by the optical head 3. The A/D conversion circuit 7 quantizes the playback signal at a sampling period, thereby converting the playback signal to a digital signal that consists of time-discrete sampled values. The equalizer 9 processes the digital signal, generating a signal having a waveform that agrees with the PR class.

The data detection circuit 10 is a signal processing circuit of the optimal row estimation method signal processing circuit for detecting data from the play-back signal waveform which is equalized to a predetermined class of a partial response, and the data detection circuit 10 specifically comprises a bit a bit detector. The decoder 11 restores a code bit row detected with the data detection circuit 10 to original user data.

The drive controller 12 is the main controller of the optical disc apparatus, and is connected to, for example, a personal computer via the interface 13 to conduct the transmission control of the record data and the play back of data. The apparatus includes a motion picture compression circuit, a motion picture expansion circuit and an error detection circuit, either not shown in the drawing. The motion picture compression circuit and the motion picture expansion circuit records and reproduces image information. The error

detection circuit detects errors from the data output from the decoder 11 and corrects the errors detected.

An output of the equalizer 8 is also input to the timing phase gradient detection circuit 18 which is a part of the timing control circuit 17. An output of this timing phase gradient detection circuit 18 is input to a VCO (a voltage control oscillator) 20 through a loop filter 19 to control the phase of the channel clock. An output of this VCO 20 is provided as a recovered channel clock, a sampling clock of the A/D conversion circuit 7 and an operation clock of the equalizer 8 and the data detection circuit (an ML detector) 10.

In an example shown in FIG. 4, the sampling frequency and the phase of the A/D conversion circuit 7 are synchronized with the play back signal by means of VCO 20. FIG. 5 is a view showing another example in which the reading clock is synchronized with the play back signal. In FIG. 5, another example is shown with respect to a portion from the VGA 6 to the data detection circuit 10.

That is, the sampling clock of the A/D conversion circuit 7 is supplied from a synthesizer 602. This sampling clock is a clock having a frequency which is somewhat higher than a channel clock period of the play back signal, and the sampling clock is not synchronized with the play back signal. An output of the equalizer

8 is shifted and output with an interpolation filter 600 by a value at which only the phase is controlled with the output of the phase calculation circuit 601.

An output of this interpolation filter 600 is input to the timing phase gradient detection circuit 18. An output of this timing phase gradient detection circuit 18 is input to a phase calculation circuit 601 via the loop filter 19, so that a shift amount of the phase in the interpolation filter 600 is controlled. A clock control 603 thins out the output of the synthesizer 602 so as to agree with the play back signal frequency on the basis of the phase control information from the phase calculation circuit 601 thereby supplying the output as an operation clock of the data detection circuit 10.

In the following explanation, it is supposed that the PR class of this PRML signal processing system is set to (1, 2, 2, 1). An operation of the timing phase gradient detection circuit 18 in this first embodiment will be explained by using FIG. 6. Thereafter, the amplitude value of the play back signal sampled in the nth sampling period is denoted by y(n).

A timing phase gradient detection circuit 18 shown in FIG. 6 is a circuit for efficiently detecting a timing phase gradient with respect to the VFO pattern shown in FIG. 7A. The VFO pattern refers to a dedicated data pattern for conducting phase matching at

a high speed, and is a repeated pattern of a single
frequency shown in FIG. 7A.

In an example of the pattern shown in FIG. 7A, the
amplitude value of the PR equalized sample point

5   becomes a repetition of values such as 1, 1, 3, 5, 5, 3,
1, 1, ··· in the case where the sampling phase matches.
However, for the simplification of the explanation on
the numeric value processing, it is supposed that the
value of $y(n)$ can assume the values of -3 through +3

10   centering on the level of 3 after the PR equalization
processing. That is, the PR equalization output level
of 1, 1, 3, 5, 5, 3, 1, 1 becomes -2, -2, 0, +2, +2.

Reference numerals 100 through 102 of FIG. 6
denote a delay circuit for outputting input sample

15   values by delaying the values in one sampling period.
In the case where the input of the delay circuit 100 is
$y(n)$, the output of each of the delay circuits 100
through 102 becomes $y(n-1)$, $y(n-2)$ and $y(n-3)$
respectively.

20   Reference numerals 109 through 111 of FIG. 6 also
denote delay circuits for outputting input sample
values by delaying the values in one sampling period.
In the case where the input of the delay circuit 109 is
$g(n)$, the output of each of the delay circuits 109

25   through 111 becomes $g(n-1)$, $g(n-2)$ and $g(n-3)$
respectively.

Here, $g(n)$ denotes a value representing a

direction of change of the amplitude value y(n) of the play back signal which is sampled in the nth sampling period. That is, the samples 150 through 152 of the waveform shown in FIG. 7F are defined as having a direction of +1 while the samples 153 through 155 are defined as having a direction of -1.

Reference numeral 108 denotes a threshold value judging circuit which judges the direction of the change of y(n) from y(n) and y(n-1) which are added with the addition circuit 103. As the threshold value η(n) for judgment, the output of the threshold value determining circuit 112 is used. In the threshold value determining circuit 112, the direction g(n-3) of the change before three sample time is input. Then in the case where the value is +1, the value of +ε is output while the value of -ε is output in the case where the value is -1.

The threshold value judging circuit 108 inputs the value of g(n-1). When g(n-3) is +1,-1 is output as g(n) in the case where y(n) + y(n-1) becomes +ε or less thereby reversing the direction of change. However, in the case where the condition is not met, the value same as g(n-1) is output. When g(n-3) is -1, +1 is output as g(n) in the case where y(n) + y(n-1) becomes -ε or less thereby reversing the direction of change. In the case where the condition is not met, the value same as g(n-1) is output.

The addition circuit 106 adds the outputs of the multiplication circuits 104 and 105 and an output of the correction value output circuit 115, generating a timing phase gradient $\Delta \tau$ 107. The multiplication circuit 104 finds the product of $y(n)$ and $g(n)$, while the multiplication circuit 105 finds the product of $y(n-3)$ and $g(n-3)$. The correction value output circuit 115 outputs the correction value for calculating the timing phase gradient $\Delta \tau$ 107 in the phase error in accordance with the comparison result of the comparator 113 and the comparison result of the comparator 114.

By using FIG. 8A and FIG. 8B and FIGS. 9A and 9B, more specific explanation will be given. In FIG. 8A, the actual sample value 204 is such that the amplitude value is $+2+\gamma$ because of the phase error $\tau 2$ with respect to the amplitude value 203 in the case where the phase of $y(n)$ matches. Furthermore, the actual sample value 202 is such that the amplitude value is set to $-2-\beta$ because of the phase error $\tau 1$ with respect to the amplitude value 201 in the case where the phase of $y(n-3)$ matches. Furthermore, since $g(n)$ is $+1$, and $g(n-3)$ is -1 as a direction of change, the output of the multiplication circuit 104 becomes

$$(+2+\gamma) \times (+1) = +2+\gamma$$

while the output of the multiplication circuit 105 becomes

$$(-2-\beta) \times (-1) = +2+\beta.$$

Furthermore, the output of the comparator 113 becomes 1 from $g(n) = g(n-1)$ because $g(n) = +1$ and $g(n-1) = +1$ are set.

The output of the comparator 114 becomes 1 from $g(n) = g(n-2)$ because $g(n) = +1$ and $g(n-2) = +1$ are set. Then, the output of the correction value output circuit 115 becomes -4.

From this fact, the timing phase gradient $\Delta \tau$ 107 in the phase error becomes $(+2+\gamma) + (+2+\beta) - 4 = \gamma + \beta$. Then, the shift from the PR equalization reference value, the sum of γ and β is output as the timing phase gradient in the phase error.

FIG. 8B is a view showing an example in which the phase error is in the reverse direction. That is, the output of the multiplication circuit 104 is

$(+2-\gamma) \times (+1) = +2-\gamma$. The output of the multiplication circuit 105 becomes

$(-2+\beta) \times (-1) = +2-\beta$.

Furthermore, the output of the comparator 113 becomes 1 from $g(n) = g(n-1)$ because $g(n) = +1$ and $g(n-1) = +1$ are set.

The output of the comparator 114 becomes 1 from $g(n) = g(n-2)$ because $g(n) = +1$ and $g(n-2) = +1$ are set, and the output of the correction value output circuit 115 becomes -4.

From this fact, the timing phase gradient $\Delta \tau$ 107 in the phase error becomes

$(+2-\gamma) + (+2-\beta) - 4 = -\gamma-\beta$.

Then the shift from the PR equalization reference value, the sum of $-\gamma$ and $-\beta$ is output as the timing phase gradient.

5   The output of the multiplication circuit 104 becomes

$(+2-\gamma) \times (-1) = -2+\gamma$

because g(n) is -1 and g(n-3) is +1 in the case where y(n) is as shown in FIG. 9A. The output of the

10   multiplication circuit 105 becomes $(-2+\beta) \times (+1) = -2+\beta$.

The output of the comparator 113 is 0, because the input g(n) (= -1) and the input g(n-1) (= +1) are not equal. The output of the comparator 114 is 0, because

15   the input g(n) (= -1) and the input g(n-2) (= +1) are not equal. Thus, the output of the correction value output circuit 115 is +4.

From this fact, the timing phase gradient $\Delta \tau$ 107 in the phase error will become $(-2+\gamma) + (-2+\beta) + 4 =$

20    $\gamma+\beta$. Then, the shift from the PR equalization reference value, the sum of γ and β is output as the timing phase gradient.

In the case where y(n) is as shown in FIG. 9B, g(n) is -1, g(n-3) is +1 so that the output of the

25   multiplication circuit 104 will be

$(-\gamma) \times (-1) = +\gamma$.

The output of the multiplication circuit 105

becomes $1 (+\beta) \times (+1) = +\beta$.

The output of the comparator 113 is 1, because the input g(n) (= -1) and the input g(n-1) (= -1) are equal. The output of the comparator 114 is 0, because the input g(n) (= -1) and the input g(n-2) (= +1) are not equal. The output of the correction value output circuit 115 is therefore 0.

From this fact, the timing phase gradient $\Delta\tau$ 107 in the phase error will be

$$(+\gamma) + (+\beta) = \gamma + \beta.$$

Then, a shift from the PR equalization reference value, the sum of γ and β is output as the timing phase gradient.

Hereinbelow, in the same manner, the shift amount from the reference value of the amplitude generated from the phase error is correctly output as the timing phase gradient.

FIG. 10 is a flowchart showing in summary the operation according to the first embodiment.

In the first embodiment, the present sample value y(n) and the sample value y(n-1) obtained at the immediately preceding sample time are applied to determine the timing phase gradient. Only the value y(n) suffices to determine the timing phase gradient correctly if the signal has no noise. If the signal has noise, the possibility of error in determining the timing phase gradient will increase. To determine the

timing phase gradient accurately, two peaks of the

signal waveform are inferred by using two sample points,

thereby determining whether the amplitude of the signal

is increasing or decreasing. Thus, above averaging,

5   minimizing the influence of the noise. It is therefore

possible to calculate the timing phase gradient at high

accuracy.

FIG. 11 is a view showing a timing phase gradient

detection circuit 18 according to a second embodiment

10   of the present invention. This timing phase gradient

detection circuit 18 is also a circuit for efficiently

detecting the timing phase gradient with respect to the

VFO pattern shown in FIG. 7A.

Reference numerals 100 through 102 of FIG. 11

15   denote a delay circuit for outputting the input sample

value 18 with a delay of one sampling period in the

same manner as FIG. 6. In the case where the input of

the delay circuit 100 is set to $y(n)$, the output of

each of the delay circuit 100 through 102 becomes

20   $y(n-1)$, $y(n-2)$ and $y(n-3)$.

Reference numerals 109 through 111 denote a delay

circuit for outputting input sample values with a delay

of one sampling period. In the case where the input of

the delay circuit 109 is set to $g(n)$, the output of

25   each of the delay circuits 109 through 111 will be

$g(n-1)$, $g(n-2)$ and $g(n-3)$.

Here, $g(n)$ is a value representing which of the

reference values +2, 0 and -2 after the PR equalization the level of the amplitude value y(n) of the play back signal which is sampled in the nth sampling period should be regarded as. When the amplitude value y(n)

5 should be regarded as +2, g(n) is +1. When the amplitude value y(n) should be regarded as 0, g(n) is 0. When the amplitude value y(n) should be regarded as -2, g(n) is -1. For the sake of convenience, this g(n) is referred to as judgment level in the following

10 explanation.

Furthermore, reference numeral 308 denotes a threshold value judging circuit for judging the value which g(n) should assume from the value of y(n). The value is separated into $\eta(n) - \delta$ which is a judgment

15 threshold value of +1 and 0, and $\eta(n) + \delta$ which is a judgment threshold value of 0 and -1 by adding $-\delta$ and $+\delta$ to the output $\eta(n)$ of the threshold value determining circuit 312 with the addition circuit 320 and 321 to be input to the threshold value judging

20 circuit 308.

This threshold value determining circuit 312 inputs the judgment level of g(n-3) before three sample time. In the case where the value is +1, the value of $+\eta$ is output. When the value is -1, the value of $-\eta$

25 is output. When the value is 0, the 0 is output.

Then, the threshold value judging circuit 308 outputs -1 as g(n) when y(n) becomes $+\varepsilon - \delta$ or less in

the case of $g(n-3) = +1$ while the threshold value

judging circuit 308 outputs 0 as $g(n)$ in the scope of

$+\varepsilon \pm \delta$. The threshold value judging circuit outputs +1

when $y(n)$ becomes $+\varepsilon+\delta$. Besides, the threshold value

5   judging circuit 308 outputs -1 as $g(n)$ when $y(n)$

becomes $-\delta$ or less in the case of $g(n-3) = 0$ while the

threshold value judging circuit 308 outputs 0 in the

scope of $+\varepsilon \pm \delta$. Then, the threshold value judging

circuit 308 outputs +1 as $g(n)$ when $y(n)$ becomes $+\delta$ or

10   more. Furthermore, the threshold value judging circuit

308 outputs +1 as $g(n)$ when $y(n)$ becomes $-\varepsilon+\delta$ or more

in the case of $g(n-3) = -1$. The threshold value

judging circuit 308 outputs 0 as $g(n)$ in the scope of

$-\varepsilon \pm \delta$ while the threshold value judging circuit 308

15   outputs -1 as $g(n)$ when $y(n)$ becomes $-\varepsilon+\delta$ or less.

The timing phase gradient $\Delta \tau$ 107 in the phase

error is output as the timing phase gradient value

selected by the selection circuit 330 on the basis of

the comparison result of the comparators 331 and 332.

20   That is, when $g(n)$ and $g(n-1)$ are not equal to each

other and the $g(n)$ and $g(n-3)$ are not equal to each

other, the value of $(-1) \times y(n) \times g(n-2) + y(n-2) \times$

$g(n)$ is output which is calculated in the addition

circuit 329 and the multiplication circuit 327 and 328.

25   Furthermore, in the case of $g(n) = g(n-1)$ and,

at the same time in the case where $g(n)$ is not equal

to $g(n-3)$, the value of

$(-1) \times y(n) \times g(n-1) + y(n-1) \times g(n)$ is output which is calculated with the addition circuit 326 and the multiplication circuits 324 and 325.

Furthermore, in the case of $g(n) = g(n-3)$, the value of $g(n-1) \times [-y(n) + y(n-3)]$ is output which is calculated in the addition circuit 322 and the multiplication circuit 323.

The operation will be explained more specifically by using FIGS. 12A and 12B and FIG. 13. In the beginning, in FIG. 12A, the actual sample value 204 is such that the amplitude value is $+2+\gamma$ because of the phase error $\tau 2$ with respect to the amplitude value 203 in the case where the phase of $y(n)$ matches. In this case, the calculation value selected as the timing phase gradient will be $(-1) \times y(n) \times g(n-2) + y(n-2) \times g(n)$ because $g(n) = +1$, $g(n-1) = 0$, $g(n-2) = -1$, and $g(n-3) = -1$.

The actual sample value 202 with respect to the amplitude value 201 in the case where the phase of $y(n-2)$ matches is such that the amplitude value is $-2+\beta$ because of the phase error $\tau 1$. As a consequence, the following result is obtained.

$$(-1) \times (+2+\gamma) \times (-1) + (-2+\beta) \times (+1) = +\gamma + \beta$$

A shift from the PR equalization reference value, namely the sum of γ and β is output as the timing phase gradient.

In FIG. 12B, it is supposed that the actual sample

value 208 is such that the amplitude value is $+2-\gamma$ because of the phase error $\tau 4$ with respect to the amplitude value 207 in the case where the phase of y(n) matches. In this case, g(n) = +1, g(n-1) = +1,

5   g(n-2) = 0, g(n-3) = -1, so that the calculation value selected as the timing phase gradient is

$$(-1) \times y(n) \times g(n-1) + y(n-1) \times g(n).$$

It is supposed that the actual sample value 206 is such that the amplitude value is $+2+\beta$ because of the

10   phase error $\tau 3$ with respect to the amplitude value 205 in the case where the phase of y(n-1) matches. As a consequence, the following result is obtained.

$$(-1) \times (+2-\gamma) \times (+1) + (+2+\beta) \times (+1) = +\gamma+\beta.$$

A shift from the PR equalization reference value,

15   namely, the sum of γ and β will be output as the timing phase gradient.

In FIG. 13, it is supposed that the actual sample value 211 is such that the amplitude value is $-\gamma$ because of the phase error $\tau 6$ with respect to the

20   amplitude value 212 in the case where the phase of y(n) matches. In this case, the calculation value selected as the timing phase gradient will be g(n-1) $\times$ [-y(n) + y(n-3)] because g(n) = 0, g(n-1) = +1, g(n-2) = +1, g(n-3) = 0 are set.

25   Suppose that the actual sample value 210 is such that the amplitude value is $+\beta$ because of the phase error of $\tau 5$ with respect to the amplitude value 209 in

the case where the phase of y(n-3) matches. Then, the following result will be obtained.

$$(+1) \times (-1) \times (-\gamma) + \beta = +\gamma + \beta$$

A shift from the PR equalization reference value, namely, the sum of γ and β is output as the timing phase gradient.

FIG. 14 is a flowchart showing in summary an operation in the second embodiment.

According to the second embodiment, the timing phase gradient can be efficiently detected by changing over the calculation method of the timing phase gradient on the basis of the history of the amplitude level which is assumed to be the assumed amplitude level.

In the first and the second embodiment described above, it is intended to determine the timing phase gradient with respect to the single frequency VFO pattern. However, even after the phase is acquired from the VFO pattern, a relatively mild phase change such as rotation drift or the like must be allowed to follow the clock. Since the user data portion is a signal generated by a random data series, the timing phase gradient cannot be obtained in a method similar to the VFO portion.

FIG. 15A through 15C are state transition diagrams for explaining how the amplitude value of the PR (1, 2, 2, 1) equalization signal changes with the data row.

FIGS. 15A through 15C are diagrams showing the cases in which the d constraint is 0, 1 and 2 respectively. The d constraint refers to a restriction such that the restriction at the modulation signal is one and a certain number of 0's must be inserted between 1 and 1 in the code after the modulation. That is, in the case of d = 0 (d constraint equal to zero), such pattern as

111001101001

may be allowed in which 1's continue after the RLL code shown in FIG. 3A.

However, in the case of d = 1, the continuation of 1 is not allowed.

Therefore, at least one 0 must be inserted between 1 and 1 as can be seen in the case of

1010001010100.

Furthermore, in the case of d = 2, two or more 0's are inserted between 1 and 1 as can be seen in

100100010010.

In a state transition diagrams shown in FIGS. 15A through 15C, Sxxx surrounded by a round represents a state by three continuous bit stream of write-data. As shown in FIG. 3C, the bit row of write-data refers to write current which is represented in a discrete time at which the RLL code row is converted into a NRZI form. In the case of d = 0, the bit stream of write-data can be changed into an arbitrary value of 0 and 1. Both

010 and 011 can be assumed as a next state of, for example, a row of 001.

RLL code   bit stream of write-data

001111   001010

001011   001101

In the case of d = 0, since at least one 0 is inserted between 1 and 1 of the RLL code, for example, only 011 can be assumed next to the bit stream of write-data of 001.

RLL code   bit stream of write-data

00101001   00110001

00100101   00111001

In the state of d = 2, since at least two 0's are inserted between 1 and 1 of the RLL code, only the state of 111 can be assumed, for example, next to the state of 011.

RLL code   bit stream of write-data

001001001 001110001

001001000 001110000

In the state transition diagrams of FIGS. 15A through 15C, numerals attached in front of the arrows denote amplitude values when the state changes. Here, the amplitude values can assume values from 0 through 6. FIGS. 16A and 16B and FIG. 17 show an amplitude pattern which can be assumed at the rise of the waveform in the case of d = 0, d = 1 and d = 2 on the basis of the state transition diagrams.

In order to determine a phase error from the PR
equalization reference value (an ideal value) and an
actual amplitude value, it is necessary to accurately
know which value the reference value at that time can
5   assume. However, if it is possible to accurately know
the value at that time, a detector at the following
step is not required. At this time, permitting that
the prediction value becomes somewhat inaccurate, there
is no other way than the adoption of a method which
10   enables avoiding an error as much as possible.
Furthermore, it is not necessary to take an error
signal from all the sample points. Detection
sensitivity may only be gained which enables the
correction of a slow phase shift.

15   In consideration of the above situation, in the
above first and the second embodiment, the phase error
information is extracted only from the portion where
the gradient of the waveform amplitude is large. A
stable clock timing recovery can be realized by
20   improving the detection precision of the timing phase
gradient by not detecting the phase error at a portion
with a small change in the amplitude which facilitates
an error in judgment of values with noises.

FIG. 18 is a view showing a timing phase gradient
25   detection circuit 18 corresponding to a random data
waveform, the view showing a third embodiment of the
present invention. In this case, in the PR class, the

row is (1, 2, 2, 1) and the d constraint is set to 1 or more. Furthermore, in the following explanation, the amplitude value can again assume a value of -3 through +3.

5   In the same manner as FIG. 6, reference numerals 100 through 102 denote a delay circuit for outputting the input sample value with a delay of one sampling period. In the case where the input of the delay circuit 100 is y(n), the output of each of the delay

10   circuits 100 through 102 becomes y(n-1), y(n-2) and y(n-3).

Furthermore, reference numerals 109 and 110 also denote a delay circuit for outputting the input sample value with a delay of one sampling period in the same

15   manner. In the case where the input of the delay circuit 109 is g(n), the output of each of the delay circuits 109 and 110 become g(n-1) and g(n-2).

Here, g(n) denotes a value showing which of the reference values +2 or more, +1, 0, -1, -2 or less

20   after the PR equalization the level of the amplitude value y(n) of the play back signal which is sampled in the nth sampling period should be regarded as. It is supposed that when the amplitude value y(n) should be regarded as +2 or more, g(n) is +1. When the amplitude

25   value y(n) should be regarded as +1, 0 and +1, g(n) is 0. When the amplitude value y(n) should be regarded as -2 or less, g(n) is -1. For the sake of convenience,

this g(n) is referred to as the judgment level
hereinafter.

In this third embodiment, an examination is made
as to whether the amplitude judgment value before and
after the sampling time of the judgment level is a
judgment value of -1 or +1 at which an amplitude
gradient becomes large. In the case where the
condition is met, the mathematical expression of the
timing phase gradient is changed over in accordance
with the condition to calculate the value. More
specifically, in the case where a combination of g(n),
g(n-1) and g(n-2) is

$$(-1 \ 0 \ 0) \qquad (+1 \ 0 \ 0)$$

$$(-1 \ 0 \ +1) \qquad (+1 \ 0 \ -1)$$

$$(0 \ 0 \ +1) \qquad (0 \ 0 \ -1),$$

the mathematical expression of the timing phase
gradient is changed over in accordance with the
condition to calculate the value.

Reference numeral 353 denotes a threshold value
judging circuit for judging the value which should be
assumed from an output value of a selection circuit 352.
The selection circuit 352 selects y(n+1) + y(n) which
is an output of the addition circuit 350 in the case
where the value of g(n-1) is 0. In the case where the
value of g(n-1) is other than 0, the selection circuit
352 selects y(n) + y(n-1) which is an output of the
addition circuit 351.

FIGS. 19A through 19F are views showing a state of the transition of the waveform in the case where the value of g(n-1) is other than 0. In the case of FIGS. 19A and 19B, both y(n-1) and y(n) are -2 or less at g(n-1) = -1. In the case of FIG. 19C, y(n-1) and y(n) are 0 and -2 respectively at g(n-1) = -1. For example, in the case where - ε = -2.5 is set at the threshold value judging circuit 353, it can be judged that the judgment level is -1 in the case of FIGS. 19A and 19B and the judgment level is 0 in the case of FIG. 19C. In the same manner, at the fall of the waveform in the same manner, it is supposed that the judgment level is +1 in the case of FIGS. 19D and 19E and the judgment level can be set to 0 in the case of FIG. 19F.

FIGS. 20A through 20F are views showing the state of the transition of the waveform in the case where the value of g(n-1) is 0. In the case of FIGS. 20A and 20B, since both y(n) and y(n+1) are +1, +2 or more at g(n-1) = 0. In the case of FIG. 20C, y(n) and y(n+1) are +1 and 0 respectively at g(n-1) = 0. For example, in the case where ε = 2.5 is set at the threshold value judging circuit 353, it can be judged that the judgment level is +1 in the case of FIGS. 20A and 20B and the judgment level is 0 in the case of FIG. 20C. In the same manner, at the fall of the waveform, it can be judged that the judgment level is set to -1 in the

case of FIGS. 20D and 20E and the judgment level is set to 0 in the case of FIG. 20F.

The timing phase gradient $\Delta \tau$ 107 in the phase error outputs the result of the addition of the output of the multiplication circuits 363, 360 and 357 with the addition circuit 356 only in the case in which g(n-1) is 0. The multiplication circuit 363 outputs the product of y(n-1) and an output of the constant determining circuit 364. The multiplication circuit 360 outputs the product of the output of the constant determining circuit 365 and the output of the addition circuit 361. The addition circuit 361 outputs the sum of y(n-2) and the result of the multiplication of the output of the constant determining circuit 365 and the constant α with the multiplication circuit 362. The multiplication circuit 357 outputs the product of the output of the constant determining circuit 366 and the output of the addition circuit 358. The addition circuit 358 outputs the sum of y(n) and the result of the multiplication of the output of the constant determining circuit 366 and the constant α with the multiplication circuit 359.

The operation will be more specifically explained by using FIGS. 21A through FIGS. 21F. Since g(n) = 0, g(n-1) = 0 and g(n-2) = -1 are set in the case of FIG. 21A, the output of the comparator 369 is 1, the output of the comparator 370 is 0, and the output of

the comparator 371 is 0, and the output of the
comparator 372 becomes 0. The timing phase gradient is
determined from the following equation.

$$y(n-1) + [y(n-2) + 1 \times \alpha] \times 1$$

5   Now, suppose that the actual sample value 204 is
such that the amplitude value is $+\gamma$ because of the
phase error $\tau 2$ with respect to the amplitude value 203
in the case where the constant $\alpha = 2$ is set, and the
phase of y(n-1) matches. Furthermore, suppose that the

10   actual sample value 202 is such that the amplitude
value is $-2+\beta$ because of the phase error $\tau 1$ with
respect to the amplitude value 201 in the case where
the phase of y(n-2) matches.

$$+\gamma + (-2+\beta+2) = +\gamma+\beta \text{ is provided.}$$

15   A shift from the PR equalization reference value,
namely the sum of γ and β is output as the timing
phase gradient.

       Since g(n) = +1, g(n-1) = 0, and g(n-2) = -1 are
set in the case of FIG. 21B, the output from the

20   comparator 369 is 1, the output of the comparator 370
is 1, the output of the comparator 371 is 0, and the
output of the comparator 372 is 0. The timing phase
gradient is determined from

$$y(n-1) + [y(n-2)+1\times\alpha] \times 1$$
$$+ [y(n)+(-1)\times\alpha] \times 1.$$

25   Now, suppose that the actual sample value 204 is
such that the amplitude value is $+\gamma$ because of the

phase error $\tau 2$ with respect to the amplitude value 203 in the case where the constant $\alpha = 2$ is set, and the phase of y(n-1) matches. Furthermore, suppose that the actual sample value 202 is such that the amplitude

5   value is $-2+\beta$ because of the phase error $\tau 1$ with respect to the amplitude value 201 in the case where the phase of y(n-2) matches. Furthermore, suppose that the actual sample value 206 is such that the amplitude value is $+2+\delta$ because of the phase error $\tau 3$ with

10   respect to the amplitude value 205 in the case where the phase of y(n) matches. Thus, the following equation is given.

$$+\gamma \; + \; (-2+\beta+2) \; + \; (+2+\delta-2) \; = \; +\gamma+\beta+\delta \; .$$

A shift from the PR equalization reference value, namely, the sum of γ, β and δ is output as the timing

15   phase gradient.

Since g(n) = +1, g(n-1) = 0, and g(n-2) = 0 are set in the case of FIG. 21C, the output of the comparator 369 is 0, the output of the comparator 370

20   is 1, and the output of the comparator 371 is 0, and the output of the comparator 372 is 0. The timing phase gradient is determined from

$$y(n-1) \; + \; [y(n)+(-1)\times\alpha] \; \times \; 1.$$

Now, suppose that the actual sample value 204 is

25   such that the amplitude value is $+\gamma$ because of the phase error $\tau 2$ with respect to the amplitude value 203 in the case where the constant $\alpha = 2$ is set, and the

phase of y(n-1) matches. Furthermore, suppose that the actual sample value 206 is such that the amplitude value is $+2+\delta$ because of the phase error $\tau 3$ with respect to the amplitude value 205 in the case where

5   the phase of y(n) matches. The following equation is given.

$$+\gamma + (+2+\delta-2) = +\gamma+\delta .$$

A shift from the PR equalization reference value, namely, the sum of γ and δ is output as the timing

10   phase gradient.

Since g(n) = 0, g(n-1) = 0, g(n-2) = +1 in the case of FIG. 21D, the output of the comparator 369 is 0, the output of the comparator 370 is 0, the output of the comparator 371 is 1, and the output of the

15   comparator 372 is 0. Thus, the timing phase gradient is determined from the following equation.

$$y(n-1) \times (-1) + [y(n-2)+(-1) \times \alpha] \times (-1).$$

Now, suppose that the actual sample value 204 is such that the amplitude value is $-\gamma$ because of the

20   phase error $\tau 2$ with respect to the amplitude value 203 in the case where the constant $\alpha = 2$ is set, and the phase of y(n-1) matches. Besides, suppose that the actual sample value 202 is such that the amplitude value is $+2-\beta$ because of the phase error $\tau 1$ with

25   respect to the amplitude value 201 in the case where the phase of y(n-2) matches. The following result is given.

$-\gamma \times (-1) + (+2-\beta-2) \times (-1) = +\gamma+\beta$.

A shift from the PR equalization reference value, namely, the sum of γ and β is output as the timing phase gradient.

Since g(n) = -1, g(n-1) = 0, and g(n-2) = +1 are given in the case of FIG. 21E, the output of the comparator 369 is 0, the output of the comparator 370 is 0 and the output of the comparator 371 is 1, and the output of the comparator 372 is 1, and the timing phase gradient is determined from

$$y(n-1) \times (-1) + [y(n-2)+(-1)\times\alpha] \times (-1)$$
$$+ [y(n)+(+1)\times\alpha] \times (-1).$$

Now, suppose that the actual sample value 204 is such that the amplitude value is $-\gamma$ because of the phase error $\tau 2$ with respect to the amplitude value 203 in the case where the constant $\alpha = 2$ is set, and the phase of y(n-1) matches. Furthermore, suppose that the actual sample value 202 is such that the amplitude value is $+2-\beta$ because of the phase error $\tau 1$ with respect to the actual value 201 in the case where the phase of y(n-2) matches. Furthermore, suppose that the actual sample value 206 is such that the amplitude value is $-2-\delta$ because of the phase error $\tau 3$ with respect to the amplitude value 205 in the case where the phase of y(n) matches. The following equation will be given.

$-\gamma \times (-1) + (+2-\beta-2) + (-2-\delta+2) \times (-1) = +\gamma+\beta+\delta$

A shift from the PR equalization reference value, namely, the sum of γ, β and δ is output as the timing phase gradient.

Since g(n) = -1, g(n-1) = 0, and g(n-2) = 0 are set in the case of FIG. 21F, the output pf the comparator 369 is 0, the output of the comparator 370 is 0, and the output of the comparator 371 is 0 and the output of the comparator 372 is 1, and the timing phase gradient is determined from

$$y(n-1) \times (-1) + [y(n)+(+1)\times\alpha] \times (-1).$$

Now, suppose that the actual sample value 204 is such that the amplitude value is $-\gamma$ because of the phase error $\tau 2$ with respect to the amplitude 203 in the case where the constant $\alpha = 2$ is set, and the phase of y(n-1) matches. Furthermore, suppose that the actual sample value 206 is such that the amplitude value is $-2-\delta$ because of the phase error $\tau 3$ with respect to the amplitude value 205 in the case where the phase of y(n) matches. The following result is given.

$$-\gamma \times (-1) + (-2-\delta+2) \times (-1) = +\gamma+\delta.$$

A shift from the PR equalization reference value, namely, the sum of γ and δ is output as the timing phase gradient.

FIG. 22 is a flowchart showing in summary an operation in the third embodiment.

In the case where the data stream is at random,

there are many cases in which the data stream to be written is at random, the amplitude value at each of the sample points of the play back waveform is such that the phase error cannot be detected in the absence of change as compared with the sample value before and after the sample value, the detection is difficult under the influence of the noise because of few change, or the detection precision is inferior.

In the third embodiment, an examination is made as to whether the amplitude judgment value before and after the sampling time on the judgment level 0 is either the judgment value of -1 or +1 at which the amplitude gradient becomes large. When the condition is met, the mathematical expression on the timing phase gradient is changed over in accordance with the condition to calculate the value. That is, since the sample point is identified at which the change in the amplitude is on the level 2 of the reference value after the PR equalization, and the timing phase gradient is calculated only at the sample point at which the influence of the noise is small by the selection of the calculation method of the timing phase gradient on the basis of the judgment value of the past and the present sample point, the detection precision/ quality is improved in the following explanation.

FIG. 23 shows the timing phase gradient detection circuit 18 corresponding to the random data system

waveform, the view showing a fourth embodiment of the present invention. Similarly to FIG. 18, PR class is (1, 2, 2, 1) and d constraint is 1 or more. Further, in the following explanation, the amplitude value is a

5 value from -3 to +3.

In the example of the previous FIG. 18, the constant α is used for canceling the PR equalization reference value for determining the timing phase gradient on the judgment level of -1 and +1. It is

10 ideal that this constant α can be actively adjust this constant α in accordance with the signal level. In the case where the constant α is set as a fixed value, the timing phase gradient circuit becomes weak with respect to the amplitude change. In the fourth

15 embodiment shown in FIG. 23, the amplitude value on the reference level is canceled without using the constant by determining the timing phase gradient using both the judgment level of +1 and -1, so that the timing phase gradient circuit is influenced by the amplitude change

20 with difficulty.

In FIG. 23, reference numerals 100 through 102 denote a delay circuit for outputting the input sample value with a delay in one sample period. In the case where the input of the delay circuit 100 is y(n), the

25 output of each of the delay circuits 100 through 102 becomes y(n-1), y(n-2) and y(n-3).

Besides, reference numerals 109 and 110 denote a

delay circuits for outputting the input sample value with a delay in one sample period. In the case where the input of the delay circuit 109 is $g(n)$, the output of each of the delay circuits 109 through 110 becomes

5      $g(n-1)$ and $g(n-2)$.

Here, $g(n)$ is a value showing which of the reference values +2 or more, +1, 0, -1, and -2 or less after the PR equalization the level of the amplitude value $y(n)$ of the play back signal which is sampled in

10   the nth sampling period should be regarded as. When the amplitude value should be regarded as +2 or more, $g(n)$ is +1. When the amplitude value should be regarded as -1, 0, and +1, $g(n)$ is 0. When the amplitude value should be regarded as -2 or less, $g(n)$

15   is -1. This $g(n)$ is referred to as a judgment level.

In this fourth embodiment, an examination is made as to whether the amplitude judgment value before and after the sampling point on the judgment level 0 is either the judgment value -1 or the judgment value +1

20   at which the amplitude gradient becomes large. In the case where the condition is met, the mathematical expression is changed over in accordance with the condition to calculate the value. More specifically, the combination of $g(n)$, $g(n-1)$, and $g(n-2)$ is

25   (-1   0   0)   (+1   0   0)

       (-1   0   +1)   (+1   0   -1)

       (0   0   +1)   (0   0   -1)

the mathematical expression of the timing phase
gradient is changed over in accordance with the
mathematical expression of the timing phase gradient to
calculate the value.

5   Reference 353 denotes the threshold value judgment
circuit for judging the value which g(n) should assume
from the output value of the selection circuit 352.
The selection circuit 352 selects y(n+1) + y(n) which
is an output of the addition circuit 350 in the case

10   where g(n-1) is 0. In the case where g(n-1) is other
than 0, y(n) + y(n-1) is selected which is an output of
the addition circuit 351. A method for determining the
judgment level in accordance with the change in the
waveform amplitude is the same as the third embodiment

15   shown in FIG. 18 using FIGS. 19A through 19F and
FIG. 20A through FIG. 20F.

The timing phase gradient $\Delta \tau$ 107 in the phase
error is such that the addition circuit 380 outputs the
addition result of the output of the multiplication

20   circuits 363 and 381 only in the case where g(n-1) is 0.
The multiplication circuit 363 outputs the product of
y(n-1) and the output of the constant determining
circuit 364. The multiplication circuit 381 outputs
the product of the output of the constant determining

25   circuit 383 and the output of the addition circuit 382.
The addition circuit 382 outputs the sum of y(n) and
y(n-2).

An operation will be explained more specifically by using FIGS. 24A through 24F. Since g(n) = 0, g(n-1) = 0, and g(n-2) = -1 are set in the case of FIG. 24A, the output of the comparator 369 is 1, the

5   output of the comparator 370 is 0, the output of the comparator 371 is 0 and the output of the comparator 372 is 0. The timing phase gradient is determined from

$$y(n-1) \times (+1) + [y(n)+y(n-2)] \times 0.$$

Suppose that an actual sample value 204 is such

10   that the amplitude value is $+\gamma$ because of the phase error $\tau 2$ with respect to the amplitude value 203 in the case where the phase of y(n-1) matches, the following result is given.

$$y(n-1) \times (+1) = +\gamma.$$

15   A shift γ from the PR equalization reference value is output as the timing phase gradient.

Since g(n) = +1, g(n-1) = 0, and g(n-2) = -1 are set in the case of FIG. 24B, the output of the comparator 369 is 1, the output of the comparator 370

20   is 1, the output of the comparator 371 is 0, and the output of the comparator 372 is 0, and the timing phase gradient is determined from the following mathematical expression.

$$y(n-1) \times (+1) + [y(n)+y(n-2)] \times 1$$

25   Suppose that the actual sample value 204 is such that the amplitude value is $+\gamma$ because of the phase error $\tau 2$ with respect to the amplitude value 203 in

the case where the phase of y(n-1) matches.

Furthermore, suppose that the actual sample value 202

is such that the amplitude value is $-2+\beta$ because of

the phase error $\tau 1$ with respect to the amplitude value

201 in the case where the phase of y(n-2) matches.

Furthermore, suppose that the actual sample value 206

is such that the amplitude value is $+2+\delta$ because of

the phase error $\tau 3$ with respect to the amplitude value

205 in the case where the phase of y(n) matches. The

following result is given.

$$+\gamma + (+2+\delta) + (-2+\beta) = +\gamma + \beta + \delta .$$

A shift from the PR equalization reference value,

namely, the sum of γ, β and δ is output as the timing

phase gradient.

Since g(n) = +1, g(n-1) = 0, and g(n-2) = 0 is set

in the case of FIG. 24C, the output of the comparator

369 is 0, the output of the comparator 370 is 1, the

output of the comparator 371 is 0, and the output of

the comparator 372 is 0. The timing phase gradient is

determined from

$$y(n-1) \times (+1) + [y(n)+y(n-2)] \times 0.$$

Suppose that the actual sample value 204 is such

that when the amplitude value is $+\gamma$ because of the

phase error $\tau 2$ with respect to the amplitude value 203

in the case where the phase of y(n-1) matches, the

following mathematical expression is given.

$$y(n-1) \times (+1) = +\gamma$$

A shift γ from the PR equalization reference value is output as the timing phase gradient.

Since $g(n) = 0$, $g(n-1) = 0$ and $g(n-2) = +1$ are set in the case of 24D, the output of the comparator 369 is 0, the output of the comparator 370 is 0, the output of the comparator 371 is 1, and the output of the comparator 372 is 0. The timing phase gradient is determined from

$$y(n-1) \times (-1) + [y(n)+y(n-2)] \times 0.$$

Suppose that the actual sample value 204 is such that when the amplitude value is $-\gamma$ because of the phase error $\tau 2$ with respect to the amplitude value 203 in the case where the phase of $y(n-1)$ matches, the following mathematical expression is given.

$$y(n-1) \times (-1) = -\gamma \times (-1) = +\gamma$$

A shift γ from the PR equalization reference value is output as the timing phase gradient.

Since $g(n) = -1$, $g(n-1) = 0$, and $g(n-2) = +1$ is set in the case of FIG. 24E, the output of the comparator 369 is 0, the output of the comparator 370 is 0, the output of the comparator 371 is 1 and the output of the comparator 372 is 1. The timing phase gradient is determined from

$$y(n-1) \times (-1) + [y(n)+y(n-2)] \times (-1).$$

Suppose that the actual sample value 204 is such that the amplitude value is $-\gamma$ because of the phase error $\tau 2$ with respect to the amplitude value 203 in

the case where the phase of y(n-1) matches.
Furthermore, suppose that the actual sample value 202
is such that the amplitude value is $+2-\beta$ because of
the phase error $\tau 1$ with respect to the amplitude value
201 in the case where the phase of y(n-2) matches.
Furthermore, suppose that the actual sample value 206
is such that the amplitude value is $-2-\delta$ because of
the phase error $\tau 3$ with respect to the amplitude value
205 in the case where the phase of y(n) matches, the
following mathematical expression is given.

$$-\gamma \times (-1) + [(-2-\delta)+(+2-\beta)] \times (-1) = +\gamma+\beta+\delta .$$

A shift from the PR equalization reference value,
namely, the sum of γ, β and δ is output as the timing
phase gradient.

Since g(n) = -1, g(n-1) = 0, and g(n-2) = 0 is set
in the case of FIG. 24F, the output of the comparator
369 is 0, the output of the comparator 370 is 0, the
output of the comparator 371 is 0 and the output of the
comparator 372 is 1. The timing phase gradient is
determined from

$$y(n-1) \times (-1) + [y(n)+y(n-2)] \times 0.$$

With respect to the amplitude value 203 in the
case where the phase of y(n-1) matches, the actual
sample value 204 is such that the amplitude value
is $-\gamma$ because of the phase error $\tau 2$, the following
mathematical expression is given.

$$y(n-1) \times (-1) = -\gamma \times (-1) = +\gamma$$

A shift γ from the PR equalization reference value is output as the timing phase gradient.

FIG. 25 is a flowchart showing in summary an operation in the fourth embodiment.

5   In the fourth embodiment, a point is selected at which the amplitude value of the play back waveform is changed by two levels on the PR reference level from the minus amplitude to the plus amplitude with respect to the central location of amplitude with two logical

10   AND circuits 384 and 385 at two continuous sample points. In the representation with the judgment value of the previous g(n), a point is selected at which a combination of g(n), g(n-1) and g(n-2) is a point of (-1 0 +1) and (+1 0 -1).

15   In this manner, when the timing phase gradient is determined by selecting a portion where the change amount of amplitude is particularly large from the waveform corresponding to the random signal stream, the change amount of amplitude by the phase shift becomes

20   large, so that the influence of the noise can be decreased while easily canceling the reference value after the PR equalization for measuring the shift with the amplitude value in the plus direction and the amplitude value in the minus direction with the result

25   that the influence of the shift from the PR reference value by the amplitude change can be exerted with greater difficulty.

A portion where the amplitude value of the play back waveform largely changes from the minus amplitude to the plus amplitude at the continuous sample point with respect to the central location of amplitude is obtained in the case where pits or a mark on the recording medium is read which is formed with the writing current pattern in which continuous 0's for at least (a partial response class interference length-1) channel clock portion and continuous 1's for at least (a partial response class interference length -1) channel clock portion are located adjacent to each other.

Specifically, since the interference length is 4 in the case of the PR class (1, 2, 2, 1), the play back waveform shown in FIG. 3E and FIG. 3F will be as follows when the writing current waveform as shown in FIG. 3C is "000111".

```
   0001221
+     1221
+       1221
 000135531
```

A large change appears in the PR equalization level 1 > level 3 > level 5 and in the play back waveform amplitude.

In the same manner, since the interference length is 3 in the case of PR (1, 2, 1) class, when the writing current waveform shown in FIG. 3C is "0011", the following result is obtained.

```
  00121
+   121
+    121
  0013431
```

5   Thus, a large change appears in the PR

equalization level 1 > level 3. Since the PR

equalization level assumes only five values from 0 to 4

in the PR (1, 2, 1) class, the change of level 1 >

level 3 becomes a maximum change that can appear.

10   In the above explanation, there has been explained

a rise edge in which the sequence of 1 continues after

the sequence of 0. However, the same thing holds true

of the fall edge in which the sequence of 0 continues

after the sequence of 1.

15   FIG. 26 is a view showing a timing phase gradient

detection circuit 18 corresponding to the random data

waveform, the view showing the fifth embodiment of the

present invention. In the same manner as FIG. 18, the

PR class is (1, 2, 2, 1) and the d constraint is 1 or

20   more. Furthermore, in the following explanation, it is

supposed that the amplitude value assumes a value from

-3 to +3.

A difference from the example in FIG. 23 described

above is that the input A of the constant determining

25   circuit 364 becomes a logical AND of the output of the

comparators 369 and 370 instead of a logical OR thereof.

In the same manner, the input B is the logical AND of

the output of the comparator 371 and the comparator 372

instead of the logical OR thereof.

With this change, in the amplitude pattern of FIG. 24A, FIG. 24C, FIG. 24D, and FIG. 24F, the timing phase gradient is not detected, and the timing phase gradient is detected only in the case of the patterns

5   in FIGS. 24B and 24E.

That is, only the point is selected in which the combination of g(n), g(n-1) and g(n-2) is (-1 0 +1) and (+1 0 -1). The timing phase gradient is detected only in the case where a portion where the amplitude value

10   of the play back waveform largely changes from the minus amplitude to the plus amplitude at the continuous sample point with respect to the central location of amplitude is obtained in the case where pits or a mark on the recording medium is read which is formed with

15   the writing current pattern in which continuous 0's for at least (a partial response class interference length-1) channel clock portion and continuous 1's for at least (a partial response class interference length-1) channel clock portion are located adjacent to each

20   other.

FIG. 27 is a flowchart showing in summary an operation according to the fifth embodiment.

According to the present invention as described in detail, even in the PR class in which the amplitude

25   levels after equalization becomes 5 through 7, the timing phase gradient can be detected from the amplitude which is sampled at the discrete time with

the result that the PLL can be driven with an output from the PR equalizer of the data detection system without providing a different circuit for the timing recovery.

5   Consequently, in the storage apparatus for recording in a high density in which the waveform interference largely appears, a synchronized clock is stably supplied to data even when the PR class is used which has a larger interference amount with the result

10   that the data detection error rate can be suppressed to a low level and a storage apparatus can be provided which can obtain a high reliability at a large capacity.

Furthermore, according to means for detecting a timing phase gradient according to the present

15   invention, the current sample value and the value before one sample time are used in the judgment of the direction of change of the amplitude with the result that the effect of removing the influence of noise is large by averaging so that the timing phase gradient

20   can be accurately calculated.

Furthermore, according to means for detecting the timing phase gradient, the timing phase gradient can be efficiently detected because the method for calculating the timing phase gradient is changed over on the basis

25   of the assumed amplitude level and the history of the assumed amplitude level.

Furthermore, according to the means for detecting

the timing phase gradient according to the present

invention, since the timing phase gradient is

determined by selecting a portion where a change amount

of amplitude is particularly large which is obtained in

5   the case where pits or a mark on the recording medium

is read which is formed with the writing current

pattern in which continuous 0's for at least (a partial

response class interference length-1) channel clock

portion and continuous 1's for at least (a partial

10   response class interference length-1) channel clock

portion are located adjacent to each other.

Consequently the change amount of amplitude by the

phase shift becomes large and the influence of noise

can be decreased while the reference value after the PR

15   equalization for measuring the shift can be easily

canceled with the amplitude value in the plus direction

and the amplitude value in the minus direction. Thus,

the influence by the shift from the PR reference value

by the change of amplitude can be exerted up with

20   greater difficulty.

In the above explanation, there has been explained

on the PR (1, 2, 2, 1) class in which when d constraint

is either 0 or 1 and the PR equalization level becomes

7 values of 0 through 6. In the PR (1, 2, 2, 1), as

25   described above, the values of 2 and 4 in the PR

equalization cannot be assumed in the case where the d

constraint is 2, so that the value of 5 can be assumed

except for the amplitude margin and the fine setting of the threshold value.

In each of the embodiments of the present invention, the amplitude values of -1 and +1 are not used in the calculation of the timing phase gradient because the level 3 constitutes the center on level 2 and level 4 on which the change of the amplitude value is small. Consequently, the method of the present invention can be also easily applied to the PR class having the value of 5 in correspondence to the case of PR (1, 2, 2, 1) when the d constraint is 2.

Additional advantages and modifications will readily occur to those skilled in the art. Therefore, the invention in its broader aspects is not limited to the specific details and representative embodiments shown and described herein. Accordingly, various modifications may be made without departing from the spirit or scope of the general inventive concept as defined by the appended claims and their equivalents.